Precoding and Distributed STBC for Asynchronous Cooperative Diversity

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***Abstract*– For mobile users without antenna arrays, transmission diversity can be achieved with cooperative space-time encoded transmissions. In this paper, a new precoding frame-based scheme with packet-wise encoding is proposed. This precoding is based on the addition of a cyclic prefix, which is overhead, so we put more information in it by implementing this cyclic prefix as a training sequence. This scheme offers us additional known symbols and enables better synchronization and (potentially) channel estimation**.

Key words : Cooperative diversity, synchronization, distributed STBC, frequency selective channels, linear block precoding, Cyclic Prefix.

1. INTRODUCTION

Multiple antennas at the receiver and the transmitter are often used to combat the effects of fading in wireless communication system. However, implementing multiple antennas at the mobile stations is impractical for most wireless applications due to the limited size of the mobile unit. So, active users pool their resources to form a virtual antenna array that realizes spatial diversity gain in a distributed fashion [3]. It is the cooperative diversity system.

Space-time coding and processing are powerful techniques for transmission diversity, among which space-time block codes (STBC) [1], [7] are especially promising because of their low computational complexity. Cooperative transmission (without STBC) has been proposed in cellular networks for cooperative diversity [6] and in sensor networks for energy efficiency and fault tolerance [5]. STBC has been naturally employed for improved bandwidth efficiency besides the targeted diversity benefits [4], [9].

So far, most existing researches on cooperative transmission assume perfect synchronization among cooperative users. Unfortunately, it is difficult, and in most cases impossible, to achieve perfect synchronization among distributed transmitters. Therefore a challenge is the lack of perfect synchronization on delay and mobility of distributed transmitters. Considering both imperfect delay synchronization and frequency selective fading, is similar to considering dispersive channels. Some scheme was proposed for synchronization [9], but the

Table 1. Transmission scheme

	n^{th} *block symbols*	$(n+1)^{th}$ *block symbols*
ant $\mathbf{tx}_1$	$\mathbf{F}_P\mathbf{S}(n)$	$\mathbf{G}_P\mathbf{S}^*(n+1)$
ant $\mathbf{tx}_2$	$\mathbf{F}_P\mathbf{S}(n+1)$	$-\mathbf{G}_P\mathbf{S}^*(n)$

block symbols length was selected appropriately such that there is no inter-frame interference, as $(N+1) \gg 2L+D$ when N is the block symbols length, L is the impulse response length of the channel and D is the maximum of the relative delay between the two cooperative mobiles.

In our work, we release this constraint on the block length, we propose a precoding based on the addition of a cyclic prefix which is implemented as a training sequence [2]. This cyclic prefix is able to cope with time dispersive channels as long as the length of their impulse response is shorter than the cyclic prefix. Moreover, due to its training sequence nature, this prefix serves for synchronization and equalization/channel estimation. We show that there is no penalty in performance or complexity and we propose a linear equalization.

2. SYSTEM MODEL

This paper focuses on the design of an asynchronous cooperative diversity system for frequency selective channels using a linear block precoding with a training sequence as cyclic prefix [2]. This is represented in Table 1 by $\mathbf{F}_P$ and $\mathbf{G}_P$. Therefore a transmitted block symbols $\mathbf{x}(n)$ is written

$$\mathbf{x}(n) = \mathbf{F}_P\mathbf{S}(n) \tag{1}$$

Where

$$\begin{aligned}\mathbf{S}(n)^T = [&s(nN), \cdots, s(nN+N-L-1)\\ &, s(nN+N-L), \cdots, s(nN+N-1)]^T\end{aligned} \tag{2}$$

which contains the training sequence $[s(nN+N-L) \cdots s(nN+N-1)]^T$, $\mathbf{G}_P$ is a matrix of $(N+L) \times (N+L)$ dimension, and $\mathbf{F}_P = \begin{bmatrix} I_{cp} \\ I_N \end{bmatrix}$ where I_N is the identity matrix, I_{cp} is a matrix contains the L latest rows on I_N, and is

Table 2. Received block symbols

	n^{th} *block sym*	$(n+1)^{th}$ *block sym*
$rx\ from\ \mathbf{tx}_1$	$r_1(n)$	$r_1(n+1)$
$rx\ from\ \mathbf{tx}_2$	$r_2(n)$	$r_2(n+1)$

received as

$$\mathbf{H}_0\mathbf{F}_P\mathbf{S}(n)+\mathbf{H}_1\mathbf{F}_P\mathbf{S}(n-1) \quad (3)$$

where $\mathbf{H}_1$ and $\mathbf{H}_0$ are respectively

$$\begin{bmatrix} 0 & \cdots & h(L) & \cdots & h(1) \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & h(L) \\ \vdots & \ddots & \dots & \ddots & \vdots \\ 0 & \dots & \dots & \dots & 0 \end{bmatrix}$$

$$\begin{bmatrix} h(0) & 0 & \cdots & \cdots & 0 \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ h(L) & \ddots & \ddots & \ddots & \vdots \\ \vdots & \ddots & \dots & \dots & 0 \\ 0 & \dots & h(L) & \dots & h(0) \end{bmatrix}.$$

The mobile target receives a summation of the signals of the two active mobiles after they travel through different paths in the channel. These channel paths induce different delays, attenuations and phase shifts to the signals cause fading in the channel. Therefore, the transmission delays and channels can be estimated efficiently from training sequences, but we propose a model with a perfect CSI in the receiver.

We define τ_1 and τ_2 respectively as the arrival time of the first and the second signals. We assume without loss of generality that $\tau_1 \le \tau_2$. The n^{th} received block symbols can be expressed as

$$r^1_{asyn}(n) = r_1(n) + r_2(n) + v(n) \quad (4)$$

where

$$\begin{aligned} r_1(n) &= H^1_0\mathbf{F}_P\mathbf{S}(n) + H^1_1\mathbf{G}_P\mathbf{S}^*(n-1) \\ r_2(n) &= \Gamma(2,1)H^2_0\mathbf{F}_P\mathbf{S}(n+1) - \Theta\mathbf{G}_P\mathbf{S}^*(n-2) + \\ &\quad \Psi(2,1)H^2_1\mathbf{F}_P\mathbf{S}(n-1) \end{aligned} \quad (5)$$

where $\Theta = [\Gamma(2,1)H^2_1 + \Psi(2,1)H^2_0]$ and $H^i_{0,1}$ are the channels between $\mathbf{tx}_i$ and the receiver. Both matrices $\Gamma(2,1)$ and $\Psi(2,1)$ with size $(N+L)\times(N+L)$ account for the asynchronism between the two signals, and they are expressed [8] respectively as

$\Gamma(2,1) = [[0_a \quad I_{ant}]; 0_P]$ and $\Psi(2,1) = [0_{a'}; [I_{post} \quad 0^T_a]]$ where $0_a = 0_{(N+L-L_\tau)\times L_\tau}$, $I_{ant} = I_{(N+L-L_\tau)\times(N+L-L_\tau)}$, $0_P = 0_{L_\tau \times N+L}$, $0_{a'} = 0_{(N+L-L_\tau)\times(N+L)}$ and $I_{post} = I_{L_\tau\times L_\tau}$. we note $L_\tau = \tau_2 - \tau_1$ the relative delay which is bounded by L.

In the same way, the $(n+1)^{th}$ received block symbols can be expressed as

$$r^1_{asyn}(n+1) = r_1(n+1) + r_2(n+1) + v(n+1) \quad (6)$$

where

$$\begin{aligned} r_1(n+1) &= H^1_0\mathbf{G}_P\mathbf{S}^*(n+1) + H^1_1\mathbf{F}_P\mathbf{S}(n) \\ r_2(n+1) &= -\Psi(2,1)H^2_1\mathbf{G}_P\mathbf{S}^*(n-2) + \Theta\mathbf{F}_P\mathbf{S}(n+1) \\ &\quad -\Gamma(2,1)H^2_0\mathbf{G}_P\mathbf{S}^*(n) \end{aligned} \quad (7)$$

and $v(n)$ and $v(n+1)$ are the corresponding noises. The receiver removes the cyclic prefix by multiplying the received signals by the matrix $R = [0_{N\times L} \quad I_N]$, it is obvious that $RH^i_1 = 0_{N\times(N+L)}$ and $R\Gamma(2,1)H^i_1 = 0_{N\times(N+L)}$ where $i \in \{1,2\}$. Due to the insertion of the cyclic prefix which contains the training sequence $R\Psi(2,1)H^2_0\mathbf{G}_P\mathbf{S}^*(n-2)$, $R\Psi(2,1)H^2_1\mathbf{F}_P\mathbf{S}(n-1)$, $R\Psi(2,1)H^2_0\mathbf{F}_P\mathbf{S}(n+1)$ and $R\Psi(2,1)H^2_1\mathbf{G}_P\mathbf{S}^*(n-2)$ are known because they have the form of $\left[0_{1\times(N-L)} \quad S_{1\times L}\right]^T$. Therefore the expressions of received signals after the cyclic prefix removal can be expressed as

$$\begin{aligned} y(n) &= RH^1_0\mathbf{F}_P\mathbf{S}(n) + R\Gamma(2,1)H^2_0\mathbf{F}_P\mathbf{S}(n+1) + Rv(n) \\ y(n+1) &= RH^1_0\mathbf{G}_P\mathbf{S}^*(n+1) - R\Gamma(2,1)H^2_0\mathbf{G}_P\mathbf{S}^*(n) \\ &\quad +Rv(n+1) \end{aligned} \quad (8)$$

so we can write $\mathbf{G}_P = \mathbf{F}_P\mathbf{G}$ where $\mathbf{G}$ is a $N\times N$ matrix, and $RH^1_0\mathbf{F}_P$ is Toeplitz and circulant, then it can be diagonalized and $RH^1_0\mathbf{F}_P = \mathbf{W}\Lambda_1\mathbf{W}^H$ where $\mathbf{W}(k,l) = \frac{1}{\sqrt{N}}exp\left(2j\pi\frac{kl}{N}\right)$ and Λ_1 is a diagonal matrix whose diagonal entries are the values of the channel transfer function:

$$\Lambda_1(k,k) = \sum_{l=0}^{L} h(l)\exp\left(-2\pi\frac{kl}{N}\right) \quad (9)$$

Now, the expressions will be

$$\begin{aligned} y(n) &= \mathbf{W}\Lambda_1\mathbf{W}^H\mathbf{S}(n) + R\Gamma(2,1)H^2_0\mathbf{F}_P\mathbf{S}(n+1) + Rv(n) \\ y^*(n+1) &= \mathbf{W}^*\Lambda^*_1\mathbf{W}^T\mathbf{G}^*_P\mathbf{S}(n+1) \\ &\quad -R\Gamma(2,1)\left(H^2_0\right)^*\mathbf{F}_P\mathbf{G}^*\mathbf{S}(n) + Rv^*(n+1) \end{aligned} \quad (10)$$

From (10), we choose $\mathbf{G}$ such that $\mathbf{W}^H = \mathbf{W}^T\mathbf{G}^*$, i.e implies only time reversal

$$\mathbf{G} = \begin{bmatrix} 1 & 0 & \cdots & \cdots & 0 \\ 0 & \ddots & \ddots & 0 & 1 \\ \vdots & \ddots & \ddots & 1 & 0 \\ \vdots & & \dots & \ddots & \vdots \\ 0 & 1 & 0 & \cdots & 0 \end{bmatrix}$$

We can remark that in the synchronous case, $\Gamma(2,1)_{syn} = \mathbf{I}_{N+L}$, and the first $(N-L)$ rows of $R\Gamma(2,1)H^2_0\mathbf{F}_P\mathbf{S}(n+1)$ and $RH^2_0\mathbf{F}_P\mathbf{S}(n+1)$ are equivalent. Then we can use the same detection techniques like in synchronous cases and we can write that $\Lambda_2 = \mathbf{W}^HR\Gamma(2,1)H^2_0\mathbf{F}_P\mathbf{W}$, which is not a diagonal matrix. We rewrite the equations in (10) we obtain

$$\begin{bmatrix} \mathbf{W}^H y(n) \\ \mathbf{W}^T y^*(n+1) \end{bmatrix} = \begin{bmatrix} \Lambda_1 & \Lambda_2 \\ -\Lambda_2^* & \Lambda_1^* \end{bmatrix} \begin{bmatrix} \mathbf{W}^H \mathbf{S}(n) \\ \mathbf{W}^H \mathbf{S}(n+1) \end{bmatrix} + \begin{bmatrix} \mathbf{W}^H Rv(n) \\ \mathbf{W}^T Rv^*(n+1) \end{bmatrix} \quad (11)$$

where

$$\mathbf{\Lambda} = \begin{bmatrix} \Lambda_1 & \Lambda_2 \\ -\Lambda_2^* & \Lambda_1^* \end{bmatrix} \quad (12)$$

2.1. Zero Forcing technique

$$\begin{bmatrix} \widehat{\mathbf{S}}(n) \\ \widehat{\mathbf{S}}(n+1) \end{bmatrix} = \begin{bmatrix} \mathbf{W} & \mathbf{0} \\ \mathbf{0} & \mathbf{W} \end{bmatrix} \left(\mathbf{\Lambda}^H \mathbf{\Lambda}\right)^{-1} \times \mathbf{\Lambda}^H \begin{bmatrix} \mathbf{W}^H y(n) \\ \mathbf{W}^T y^*(n+1) \end{bmatrix} \quad (13)$$

2.2. MMSE technique

$$\begin{bmatrix} \widehat{\mathbf{S}}(n) \\ \widehat{\mathbf{S}}(n+1) \end{bmatrix} = \begin{bmatrix} \mathbf{W} & \mathbf{0} \\ \mathbf{0} & \mathbf{W} \end{bmatrix} \left(\mathbf{\Lambda}^H \mathbf{\Lambda} + \frac{\sigma_n^2}{\sigma_s^2} \mathbf{I}_{2\times N}\right)^{-1} \times \mathbf{\Lambda}^H \begin{bmatrix} \mathbf{W}^H y(n) \\ \mathbf{W}^T y^*(n+1) \end{bmatrix} \quad (14)$$

where $\frac{\sigma_s^2}{\sigma_n^2}$ is the signal to noise ratio. Finally, we can notice the ease of detection, although the relative delay can be as long as the training sequence period. The asynchronism effects will be evaluated by simulations.

3. SIMULATIONS

We consider a multipath channel which is frequency selective, we use a normalized power delay profile with an RMS delay spread of 0.9221 μsec, then a coherence bandwidth of 216.9 KHz. and we choose the bandwidth of the system equal to 2 MHz. We consider a block fading. We suppose that the system is radiation power limited, then the energy allocated to each symbol should be halved in order to have the same total radiated power from two transmit antennas. The relative delay is randomly generated.

Since the ZF detector (13) is linear and immune to ISI, in the presence of Additive White Gaussian Noise (AWGN) with zero mean ans variance σ_n^2, its performance in the synchronous case can be derived in closed form. Then for each realization of the channel impulse responses, the BER expressions on the ith symbol of the nth block, achievable with the symbol-by-symbol detector and assuming a BPSK constellation, for the one transmit antenna and the two transmit antennas cases are respectively [10]

$$\frac{1}{2N} \sum_{k=1}^{N} erfc\left(\sqrt{\frac{\sigma_s^2}{\sigma_n^2 B(k,k)}}\right) \quad (15)$$

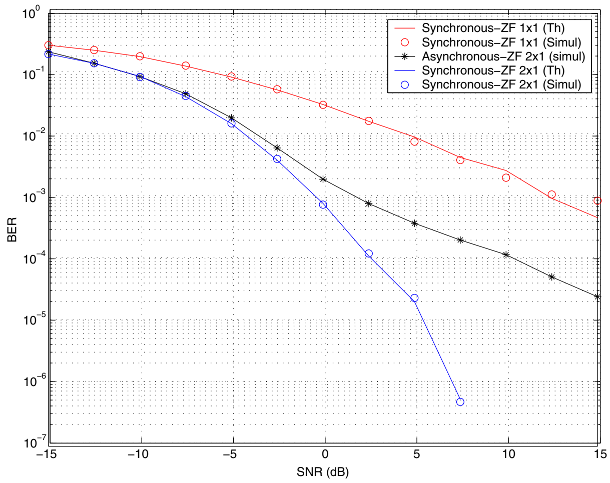

Figure. 1. Performance comparison of the precoding-distributed STBC with ZF detection

where $B = W\Lambda^{-1}\left(\Lambda^{-1}\right)^H W^H$

$$\frac{1}{2N} \sum_{k=1}^{N} erfc\left(\sqrt{\frac{\sigma_s^2}{2\sigma_n^2 M(k,k)}}\right) \quad (16)$$

where $M = W\left(\Lambda_1 \Lambda_1^H + \Lambda_2 \Lambda_2^H\right)^{-1} W^H$.

We have checked our theoretical results with simulation in

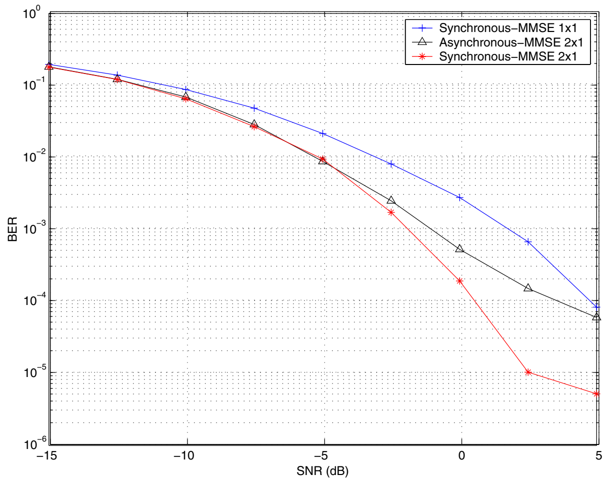

Figure. 2. Performance comparison of the precoding-distributed STBC with MMSE detection

the synchronous case to which we add the simulation of asynchronous system with ZF and MMSE detector, see [Fig 1 and

2]. We report the BER averaged over 1000 independent channel realizations, as function of the SNR. The channel is FIR of order 22 and the cyclic prefix CP of length $L = 64$ contains the training sequence. The frame contains 320 symbols of which 192 data information symbols. The relative delay L_τ is up to 10. In Fig(2), we keep the diversity in the asynchronous scheme.
We remark the applicability of the scheme with the MMSE even in the worst of the cases, when L_τ is up to $64(= L)$.[see Fig3].
With a Linear Time Invariant channel, both detectors can be implemented in $\mathcal{O}(N^2) + \mathcal{O}(N \log N)$ operations. The results prove to be good and the use of precoding reduces the complexity than [9] which presents a linear system resolution with at least $\mathcal{O}(N^3)$ operations, making it impractical for large N. In [9] the frame contains 500 symbols, the channel is only 4 taps, and the relative delay is bounded by $L_\tau \leq 10$.

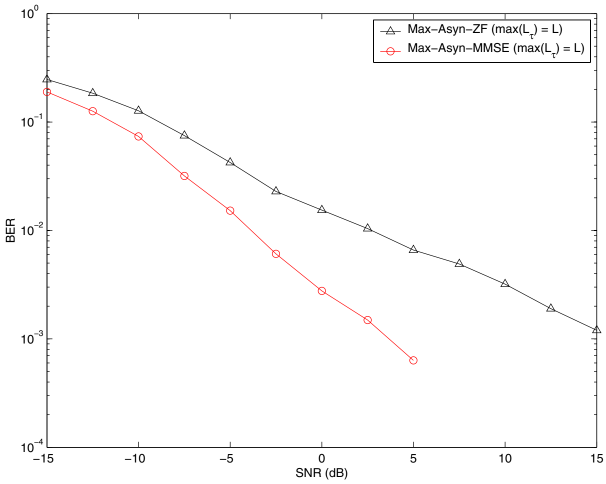

Figure. 3. Comparison between ZF and MMSE for the worst asynchronism case

4. CONCLUSION

In this paper, we proposed a new precoding frame-based scheme with packet-wise encoding, which is implemented as $CP-training$ sequence. The new scheme tolerates a relative delay as large as the cyclic prefix length. Full transmission diversity can be achieved with linear computational complexity. Our STBC-CP scheme can be extended to the number N_t of transmit antennas and to any number of receive antennas [7]. We enhance the spectral efficiency by diminishing the length of the cyclic prefix, we can reach 0.78 when 32 and 288 are respectively the length of the cyclic prefix and the frame.

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